Diversity and Multiplexing Techniques of 802.11n WLAN

Maciej Krasicki

Abstract—This paper is devoted to analyze an improvement in the performance of WLAN (Wireless Local Area Network) systems introduced by space and space-time diversity, as well as spatial multiplexing. These MIMO (Multiple-Input Multiple-Output) techniques are approved in the latest 802.11n specification. In order to perform the experiment, a Matlab application that simulates WLAN physical layer has been developed.

Index Terms—Signal processing, MIMO systems, diversity schemes, coding, modulation.

I. INTRODUCTION

COMMON WLAN standards defined by IEEE operate in the ISM (Industrial, Scientific, Medical) bands, i.e. 2.4 GHz and 5.2 GHz. OFDM (Orthogonal Frequency Division Multiplexing) is applied to overcome intersignal interference (ISI). The transmission runs in a frame mode. Numerous Modulation and Coding Schemes (MCS) are provided, which are switched by the transmitter adaptively, according to the channel condition.

The new specification of WLAN systems [1] has introduced many techniques to improve data rate in the physical layer. Apart from modification of the OFDM symbol (52 subcarriers dedicated for data transmission instead of 48 in 802.11a/g, shorter guard interval), two groups of methods can be distinguished: with backward signaling and without it. The first group comprises beamforming, i.e. based on knowledge of the channel state, the transmitter forms the signals in such a way that their performance at the receiver’s input is optimized. These methods are not considered in the paper, which focuses on the space and space-time diversity techniques, instead. Spatial multiplexing is also addressed.

Some results of multi-antenna OFDM systems performance have been delivered in a few articles, e.g. [2], [3]. They can be treated as a reference to the present work to verify the accuracy of the simulation Matlab code developed by the author.

The article is organized as follows: Section 2 reviews space and space-time diversity techniques, while Section 3 refers to spatial multiplexing. The simulation results are presented in Section 4. Finally, Section 5 concludes the work.

II. SPACE AND SPACE-TIME DIVERSITY SCHEMES

The aim of space and space-time diversity is to improve radio link quality, by means of MIMO technology. In the first place, the systems with only receive diversity will be considered. Afterwards, a smart idea of Space-Time Block Coding (STBC) [4], which is proposed by 802.11n specification, will be examined. A general model of the transmitter and the receiver of a system employing space (space-time) diversity is shown in Fig. 1. At the transmitter, adjacent data bits are encoded by a convolutional encoder. Consecutive codewords are distributed among adjacent subcarriers according to the block interleaving rule, after which they are mapped onto signals $C_k(p)$, where $k$ is the number of subcarrier and $p$ denotes the number of OFDM symbol.

The STBC encoder (if implemented) takes the consecutive signals $C_k(p)$ and $C_k(p+1)$, occupying a given subcarrier $k$, which fall to the $p$-th and the $(p+1)$-th OFDM symbols, and creates their modified copies. All the signals are transmitted according to the orthogonal Alamouti scheme [4], i.e. the first antenna transmits $C_{k1}(p) = C_k(p)$ and $C_{k1}(p+1) = -C_k^*(p+1)$ on the $p$-th and the $(p+1)$-th OFDM symbol, respectively. Simultaneously, the second antenna transmits $C_{k2}(p) = C_k(p+1)$ and $C_{k2}(p+1) = C_k^*(p)$. The signals to be transmitted via the second antenna are cyclically rotated, according to 802.11n specification, but this operation does not result in further diversity gain.

If space-time diversity is not implemented, STBC block is “transparent”, i.e. $C_{k1}(p) = C_k(p)$, $C_{k1}(p+1) = C_k(p+1)$, etc. In this case only one stream is transmitted.

Next, OFDM is performed by means of Inverse Fast Fourier Transformation (IFFT). Finally, Cyclic Prefix is added to avoid inter-signal interference. In a real system Digital/Analog conversion and carrier modulation should be done before the signals are transmitted. These steps can be omitted in simulations since the transmission in a baseband channel is considered.

At the receiver, after Cyclic Prefix removal (CPR) and
OFDM demodulation (FFT algorithm), each subchannel in the frequency domain is ideally estimated, i.e. the frequency responses $H_{knm}$ of the subchannel between the $n$th transmit and the $n$-th receive antenna at the $k$-th subcarrier are calculated for all $m, n, k$. If the frequency response does not vary while a data frame is transmitted, the time index $p$ can be omitted. The signal received from the $n$-th antenna at the $k$-th subcarrier in the $p$-th OFDM symbol is

$$R_{kn}(p) = \sum_{m} H_{knm} C_{km}(p) + \eta_{kn}(p),$$  \hspace{1cm} (1)$$

where $C_{km}(p)$ is a signal transmitted from the $m$-th antenna at the $k$-th subcarrier in the $p$-th OFDM symbol, $\eta_{kn}$ is a component representing additive noise. The diversity combiner computes estimates $\tilde{C}_k(p)$ of the transmitted signals, in a way depending on the employed diversity scheme. It delivers estimates $\tilde{H}_k$ of the effective channel frequency response to the Maximum Likelihood detector, which makes decisions about the transmitted codewords. Finally, the deinterleaved bits are passed to the Viterbi decoder.

A. Receive Diversity

The following diversity algorithms are to be examined: Antenna Selection, Subcarrier Selection, Equal Gain Combining (EGC) and Maximal Ratio Combining (MRC). Since only one transmit and two receive antennas are used, let us denote $\mathbf{H}_n = [H_{n11} \ldots H_{n64}]^T$, $R_n(p) = [R_{1n}(p) \ldots R_{64n}(p)]^T$, $\mathbf{C}(p) = [\mathbf{C}_1(p) \ldots \mathbf{C}_{64}(p)]^T$, and finally $\mathbf{H} = [\mathbf{H}_1 \ldots \mathbf{H}_{64}]^T$.

1) Antenna Selection: The diversity combiner chooses a signal with higher average power from the signals received by adjacent antennas. Thus $\mathbf{C}(p) = \mathbf{R}_1(p)$ and $\mathbf{H} = \mathbf{H}_1$ if $\sum_k |H_{kl1}|^2 > \sum_k |H_{kl2}|^2$. Otherwise, $\mathbf{C}(p) = \mathbf{R}_2(p)$ and $\mathbf{H} = \mathbf{H}_2$. It is noticeable that the comparison of average power is executed only once per frame due to the assumption of channel stationarity.

2) Subcarrier Selection: The choice of antenna is made separately for each subcarrier $k$, depending on the magnitude response. That is $\tilde{C}_k(p) = R_{k1}(p)$ and $\tilde{H}_k = H_{k11}$ if $|H_{kl1}| > |H_{kl2}|$. Otherwise $\tilde{C}_k(p) = R_{k2}(p)$ and $\tilde{H}_k = H_{k21}$.

3) Equal Gain Combining (EGC): The signals from both receive antennas are exploited, i.e. they are added after the compensation of phase offsets:

$$\tilde{C}_k(p) = R_{k1}(p)e^{-j \arg(H_{k11})} + R_{k2}(p)e^{-j \arg(H_{k21})}.$$ 

Consequently $\tilde{H}_k = |H_{k11}| + |H_{k21}|$. The same operation runs for each subcarrier.

4) Maximal Ratio Combining (MRC): This technique is very similar to EGC. The only modification is that the signals from both antennas are weighted according to their power. Hence, the estimated transmitted signals are computed as $\tilde{C}_k(p) = R_{k1}(p)H_{k11} + R_{k2}(p)H_{k21}$, while the estimates of the effective channel response can be written as $\tilde{H}_k = |H_{k11}|^2 + |H_{k21}|^2$.

B. Space-Time Block Codes

In case of space-time coding, the diversity combiner computes the estimates of transmitted signals again. It is done according to the following routine. The signals received by adjacent antennas in consecutive timeslots $p$ and $p+1$ can be written as:

$$R_{k1}(p) = H_{k11}C_k(p) + H_{k12}C_k(p+1)e^{-j\theta} + \eta_{k1}(p)$$
$$R_{k1}(p+1) = -H_{k11}C_k^*(p+1) + H_{k12}C_k^*(p)e^{-j\theta} + \eta_{k1}(p+1)$$
$$R_{k2}(p) = H_{k21}C_k(p) + H_{k22}C_k(p+1)e^{-j\theta} + \eta_{k2}(p)$$
$$R_{k2}(p+1) = -H_{k21}C_k^*(p+1) + H_{k22}C_k^*(p)e^{-j\theta} + \eta_{k2}(p+1).$$

The factor denoted by $e^{-j\theta}$ represents the phase rotation, required by 802.11n specification, which has to be compensated at the receiver. The author of this paper proposes to modify the original routine of diversity combiner [4] to mitigate the effect of cyclic rotation, introduced by the transmitter:

$$\tilde{C}_k(p) = R_{k1}(p)H_{k11} + H_{k12}(R_{k1}(p)+H_{k12}(p+1)e^{j\theta})^* + H_{k21}(R_{k2}(p)+H_{k22}(p+1)e^{j\theta})^*$$
$$\tilde{C}_k(p+1) = R_{k1}(p+1)e^{j\theta} - H_{k11}(R_{k1}(p+1))^* + H_{k22}(R_{k2}(p+1)e^{j\theta} - H_{k21}(R_{k1}(p+1))^*.$$ 

It can be proved that each of these combined signals relates to a single transmitted signal. In case of the $2 \times 1$ STBC system, the components associated with signals received from the second antenna should be omitted in (3).

III. SPATIAL MULTIPLEXING

Spatial multiplexing offers higher data rate than any of diversity techniques analyzed above. The transmitter and receiver structures are shown in Fig. 2. Consecutive bits outgoing from the encoder are distributed among different space streams and are subject to constellation mapping, cyclic shift and IFFT.

As two independent signals are transmitted simultaneously through different antennas, they interfere with one another at the input of the receiver. To overcome this disadvantage, a simple Zero Forcing combiner is employed, which evaluates the estimates of signals $\tilde{C}_k(p) = [C_{k1}(p) \ldots C_{km}(p)]^T$, transmitted from antennas $1 \ldots m$ at the $k$-th subcarrier. Let us...
Fig. 3. Average power delay profile

denote $\mathcal{R}_k(p) = [R_{k1}(p) \ldots R_{kn}(p)]^T$ and

$$
\mathcal{H}_k = \begin{bmatrix}
H_{k11} & \cdots & H_{k1m} \\
\vdots & \ddots & \vdots \\
H_{kn1} & \cdots & H_{knm}
\end{bmatrix}
$$

It is noticeable that $\mathcal{R}_k(p) = \mathcal{H}_k \mathcal{C}_k(p) + \eta_k(p)$. To recover the transmitted signals, $\mathcal{R}_k(p)$ is multiplied by the inverse channel matrix $\mathcal{H}_k^{-1}$. Note that in case of spatial multiplexing, there is no need to balance the cyclic shifts, which can be handled as if they were introduced by the channel. After ZF combining, the signals are demapped and deinterleaved, as for diversity techniques, but separately in different space streams. Finally, demultiplexed bits undergo convolutional decoding.

IV. SIMULATION RESULTS

A. Simulation setup

Timing-related properties are inherited from 802.11n specification. Transmission runs in the 20 MHz bandwidth mode, 52 subcarriers are dedicated for data transmission, 4 of them are assigned to pilot signals. The convolutional encoder characterized by $[171\,133]_{OCT}$ generator polynomials is employed (resultant data rate is 1/2). Two modulation schemes are considered: QPSK and 16-QAM. An average total power is 1 W. It is independent of the number of transmit antennas, for a fair comparison.

A subchannel between each transmit and each receive antenna is simulated according to the 11-tap exponential model (see e.g. [5]) with the root-mean-square delay spread $t_{\text{rms}}$ of 92.435 ns. The average power delay profile of the assumed subchannel is shown in Fig. 3. Randomly generated fading coefficients are normalized to achieve unitary average signal power at the input of each receive antenna. The assumed subchannel model is similar to ETSI B [6] in terms of the rms delay spread but much easier to simulate.

The Doppler effect, a result of evolving channel state, has been neglected. To justify this approach, let us assume the terminal speed $v = 3$ km/h and the carrier frequency $f_c = 2.45$ GHz. Then, the maximum Doppler shift is $f_{D_{\text{max}}} = v f_c / c \approx 6.8$ Hz ($c$ is the speed of light). In the auto-regressive channel model (see e.g. [7]), the time-domain channel response of the $j$-th tap of the subchannel at discrete time $t + iT_s$ is

$$
g_j(t + iT_s) = \alpha_i g_j(t) + w_j(t + iT_s)
$$

where $\alpha_i = E \left( g_j(t) g_j^* (t + iT_s) \right) = J_0 (2 \pi f D_{\text{max}} T_s)$, $E(\bullet)$ denotes the expected value, $J_0(\bullet)$ is the zeroth-order Bessel function of the first kind, $w_j(t + iT_s)$ is an independent complex Gaussian random variable with zero mean and variance $\sigma^2 = 1 - \alpha_i^2$. $T_s$ is the sample time. As the worst case, 4096 information bytes per frame are to be transmitted in mode 1 (BPSK) without spatial multiplexing. The resultant number of the OFDM symbols is 1261, that gives 100880 samples in time domain (including the cyclic prefix). The autocorrelation value of tap responses falling to a frame declines only from 1 to 0.988. It proves that the Doppler effect can be neglected. Assuming that each frame is transmitted in different channel condition due to random channel access, fading coefficients can be generated independently for each frame.

B. Results

First, let us consider Single-Input Single-Output systems ($\text{MCS} \in \{1, 3\}$). The BER curves for 16-QAM and QPSK are presented in Fig. 4.a and Fig. 5.a, respectively, with thin solid lines. The analyzed curves are asymptotically parallel since both systems have the same number of antennas. The higher modulation order, i.e. the number of bits mapped onto one constellation point, the worse BER performance. But it does not mean that 16-QAM is worse than QPSK in any case. To make the comparison fair, higher data rate of the former should be taken into account. Moreover, any erroneously decoded bit is the cause of frame retransmission. Therefore, $\text{Throughput} = R (1 - \text{FER})$, where $R$ denotes the data rate and FER is the Frame Error Rate, is a more accurate measure of the link quality. Charts displaying the throughput are shown in Fig. 4.b and Fig. 5.b, respectively. The notation of particular curves is the same as before. It turns out that the 16-QAM system outperforms the QPSK one for SNRs $> 19$ dB, giving higher throughput.

The receive diversity schemes reviewed in Section 2 have been examined for 16-QAM and QPSK. It is noticeable that Antenna Selection is rather an inferior technique, while the others significantly improve data link quality (higher slope of BER curve, diversity gain of about 10 dB around the BER of $10^{-6}$). The difference in BER between particular algorithms is negligible, but only EGC and MRC are comparable with each other in the throughput, so there is a suggestion to employ Equal Gain Combining, due to its easier implementation.

For comparison, the $2 \times 1$ system with Space-Time Block Code has been analyzed. The BER and throughput curves are shifted right by about 3 dB in comparison with EGC. It is justified by the fact that the total transmitted power is normalized. In consequence, the power per receive antenna is still the same, and hence the systems with multiplied receive antennas perform better. Therefore, receive diversity techniques are more advantageous than Space-Time Block Coding, the more so as they are easier to implement. Nevertheless, space-time
codes are still useful to build a system with diversity only at one (Access Point’s) side.

The performance of the $2 \times 2$ STBC 16-QAM system has been examined, too. It appears to be much better than any $1 \times 2$ or $2 \times 1$ system since the signals are transmitted through 4 independent subchannels (additive noise varies from one time sample to another). SNR gain of about 15 dB around the BER of $10^{-6}$ is observed in comparison with the SISO system.

Finally, the advantages of spatial multiplexing have been analyzed. The BER and throughput curves of $2 \times 2$ and $4 \times 4$ 16-QAM $(MCS \in \{9, 27\})$ as well as $2 \times 2$ QPSK $(MCS = 11)$ systems are shown in Fig. 4 and Fig. 5, respectively. As it can be noticed, the multiplexed systems offer the same BER performance as $1 \times 1$ ones, asymptotically. Nevertheless, at low SNRs the signal detection is destroyed by the additive noise gained by the ZF combiner. In the region of high SNRs, the throughput is higher than for the $1 \times 1$ system, proportionally to the number of space streams on both sides of the system.

V. CONCLUSIONS

In this paper some transmit and receive diversity algorithms, approved by 802.11n specification, have been analyzed. These MIMO techniques have appeared to be powerful tools to enhance data rate regardless of the channel state. Thanks to $2 \times 1$ Space-Time Block Codes, the system with antennas doubled only on the Access Point’s side can improve the link quality in both directions. Spatial multiplexing enhances the throughput but it fails in case of poor channel condition, which is caused by the ZF operation. To overcome this disadvantage, other algorithms, such as Minimum Mean Square Error (MMSE) [8] and Successive Interference Cancellation (e.g. [9]), should be examined in the future.

The conclusions the author arrived at agree with earlier works related to MIMO-OFDM schemes. The simulation Matlab code passed the validation test and, therefore, it can be used in further research.

REFERENCES

[6] BRAN TS 101 475 v1.2.2 BRAN; HIPERLAN Type 2; Physical (PHY) layer.
Maciej Krasicki received the M.S. degree in Electronics and Telecommunications from Poznan University of Technology, Poland, in 2006. Since then he has been working towards the Ph.D. degree. His dissertation work concerns a new (‘boosted’) space-time diversity scheme, designed to support iterative decoding at the receiver of WLAN systems. His Ph.D. defense took place in 2010. From 2009 he has been with the Faculty of Electronics and Telecommunications, Poznan University of Technology, as a Research Assistant. His research interests include multi-antenna transmission, space-time coding and iterative signal processing. He has published several papers in journals (e.g., Electronics Letters) and conference proceedings.